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15. SUBJECT TERMS
Code-channel allocation, code-division multipleaccess, cognitive radio, power allocation, semidefinite programming, signal-to-interference-plus-noise ratio.

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Cognitive Code-Division Channelization

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Abstract—We consider the problem of simultaneous power and code-channel allocation for a secondary transmitter/receiver pair coexisting with a primary code-division multiple-access (CDMA) system. Our objective is to find the optimum transmitting power and code sequence of the secondary channel that maximize the signal-to-interference-plus-noise ratio (SINR) at the output of the maximum SINR linear receiver, while at the same time the SINR of all primary channels at the output of their max-SINR receiver is maintained above a certain threshold. This is a non-convex NP-hard optimization problem. We propose a novel feasible suboptimum solution using semidefinite programming. Simulation studies illustrate the theoretical developments.

Index Terms—Code-channel allocation, code-division multiple-access, cognitive radio, power allocation, semidefinite programming, signal-to-interference-plus-noise ratio.

I. INTRODUCTION

RECENT experimental studies [1] demonstrated that much of the licensed radio spectrum experiences low utilization. Cognitive radio (CR) [2] emerges as a promising technology to improve spectrum utilization by allowing secondary users/networks to share spectrum licensed by primary users. As licensees, the primary users are to have guaranteed access to the spectrum [3]. Therefore, the underlying challenge of CR technology is to ensure the Quality-of-Service (QoS) requirements of the primary users and, simultaneously, maximize in a best-effort context the QoS of the secondary users [4]-[7].

Herein, we are particularly interested in cognitive radio built around a code-division multiple-access (CDMA) primary system. In contrast to frequency or time division operation where cognitive secondary users may transmit opportunistically in sensed spectrum holes/void only, cognitive code-division users may in principle operate in parallel in frequency and time to a primary system as long as the induced spread-spectrum interference remains below a pre-defined acceptable threshold[1]. Power control for cognitive code-division systems was considered in [9] under an "interference temperature" constraint (total secondary user disturbance power over primary band). No optimization was carried out with respect to the code channels (signatures) of the secondary users in [9]. In contrast, in [10] a secondary code assignment scheme was presented to minimize the mean-square crosscorrelation of the secondary code with the primary received signal. Extension to multiple secondary users was also considered in the form of iterative secondary code set construction. Under interference minimizing code assignments, bit rate and spreading factor adjustments for a secondary CDMA system were considered in [11]. Interesting work outside the framework of CDMA CR in the form of joint beamforming and power allocation algorithms was reported in [12], [13], while auction mechanisms for power control were presented in [14].

In this paper, we consider the opportunity to establish a secondary code-division link coexisting with a primary CDMA system. In particular, we study for the first time the problem of designing a joint power and code-channel allocation scheme for the secondary link that maximizes the output Signal-to-Interference-plus-Noise-Ratio (SINR) of the maximum-SINR linear receiver filter under SINR constraints on all primary users[2] and a peak transmission power constraint on the secondary user. We recognize that, regretfully, this key CR CDMA formulation is a non-convex NP-hard problem. Yet, using semidefinite programming methodology we are able to develop a novel, realizable suboptimum solution with excellent cognitive system performance characteristics as demonstrated by simulation studies included in this paper.

The rest of the paper is organized as follows. Section II is devoted to the description of the CR CDMA system model and our formulation of the optimization problem. In Section III, we present in detail our proposed power and code-channel allocation solution. The performance of the proposed scheme is evaluated through simulations in Section IV. A few concluding remarks are drawn in Section V.

II. SYSTEM MODEL AND PROBLEM STATEMENT

We consider a primary CDMA system with processing gain (code sequence length) L, K primary transmitters $PT_i, i = 1, 2, \ldots, K$, and a primary receiver PR (for example, K uplink transmissions by users $PT_i, i = 1, 2, \ldots, K$, to base station PR). We also consider a potential concurrent secondary code-division link in the spectrum band of the primary system between a secondary transmitter ST and receiver SR (Fig. 1). All signals, primary and secondary, are supposed to propagate over flat-fading channels and experience additive

[1]While early standardization and regulation discussions have begun [8], no conclusive "interference temperature" rules and agreements have been reached yet.

[2]The given SINR constraints on the primary signals may be drawn/derived from preset QoS requirements for the primary system.
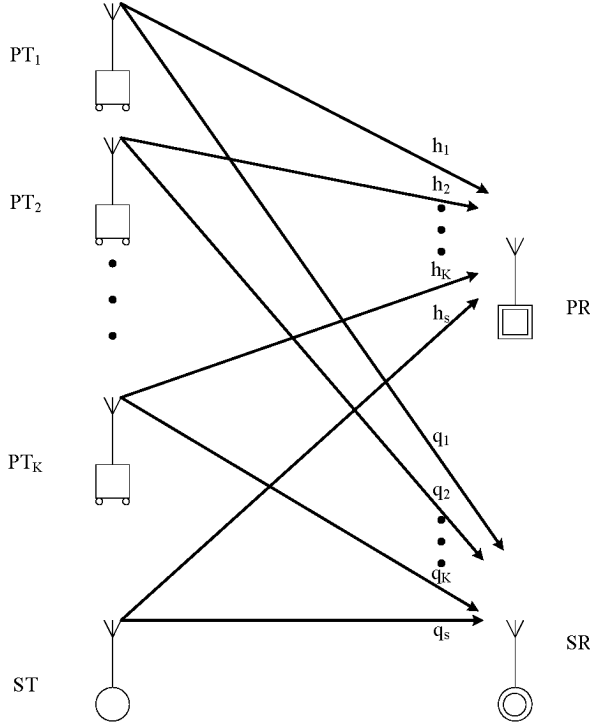
Fig. 1. Primary/secondary CDMA system model of K primary transmitters $PT_i, i = 1, 2, \ldots, K$, a primary receiver PR, and a secondary transmitter-receiver pair ST, SR. All paths $h_1, \ldots, h_K, q_1, \ldots, q_K, h_s, q_s$ exhibit independent (quasi-static) Rayleigh fading.

white Gaussian noise. We denote by h_i, q_i, $i = 1, 2, \ldots, K$, the path coefficients from PT_i to PR and SR, respectively. The path coefficients from ST to PR, SR are denoted by h_s and q_s, respectively. All path coefficients are modeled as Rayleigh distributed random variables that are independent across user signals and remain constant during several symbol intervals (quasi-static fading).

After carrier demodulation, chip-matched filtering and sampling at the chip rate over the duration of a symbol (bit) period of L chips, the received signal at the primary receiver PR can be represented as

$$\mathbf{r} = \sum_{i=1}^{K} \sqrt{E_i} h_i \mathbf{s}_i b_i + \sqrt{E_s} h_s \mathbf{s}_s b_s + \mathbf{n}_p, \quad (1)$$

while the secondary signal received by SR is

$$\mathbf{y} = \sum_{i=1}^{K} \sqrt{E_i} q_i \mathbf{s}_i b_i + \sqrt{E_s} q_s \mathbf{s}_s b_s + \mathbf{n}_s \quad (2)$$

where $E_i > 0$, $b_i \in \{\pm 1\}$, and $\mathbf{s}_i \in \mathbb{R}^L$, $\|\mathbf{s}_i\| = 1$, denote bit energy, information bit, and normalized signature vector of primary user i, $i = 1, 2, \ldots, K$, respectively; $E_s > 0, b_s \in \{\pm 1\}$, and $\mathbf{s}_s \in \mathbb{R}^L$, $\|\mathbf{s}_s\| = 1$, denote the bit energy, information bit, and normalized signature vector, respectively, of the secondary transmitter ST; $\mathbf{n}_p$ and $\mathbf{n}_s$ represent additive white Gaussian noise (AWGN) at PR and SR, correspondingly, independent from each other with $\mathbf{0}$ mean and autocovariance matrix $\sigma^2 \mathbf{I}$.

The linear filters at the primary and secondary receivers that exhibit maximum output SINR [15] can be found to be

$$\mathbf{w}_{maxSINR,i} = c_1 \mathbf{R}_p^{-1} \mathbf{s}_i, \ i = 1, 2, \ldots, K,$$
$$\mathbf{w}_{maxSINR,s} = c_2 \mathbf{R}_s^{-1} \mathbf{s}_s,$$

where $\mathbf{R}_p = E\{\mathbf{r r}^T\}$, $\mathbf{R}_s = E\{\mathbf{y y}^T\}$, $c_1, c_2 > 0$ ($E\{\cdot\}$ denotes statistical expectation and T is the transpose operator). The output SINR at PR with respect to the signal transmitted by PT_i, $SINR_i$, is given in (3) followed in (4) by the output SINR at SR, $SINR_s$, where $\mathbf{R}_{p/i}$ and $\mathbf{R}_{s/s}$ are the "exclude i" and "exclude s" input data autocorrelation matrices at PR and SR, respectively, defined by

$$\mathbf{R}_{p/i} \triangleq \sum_{k=1, k \neq i}^{K} E_k h_k^2 \mathbf{s}_k \mathbf{s}_k^T + E_s h_s^2 \mathbf{s}_s \mathbf{s}_s^T + \sigma^2 \mathbf{I},$$

$$\mathbf{R}_{s/s} \triangleq \sum_{k=1}^{K} E_k q_k^2 \mathbf{s}_k \mathbf{s}_k^T + \sigma^2 \mathbf{I}.$$

In our cognitive radio setup, the secondary transmitter has to guarantee the SINR QoS of all primary users. In this spirit, our objective is to find the transmission bit energy E_s and the real-valued normalized signature vector $\mathbf{s}_s$ that maximize $SINR_s$ under the constraints that $SINR_i$, $i = 1, 2, \ldots, K$, are all above a certain threshold $\alpha > 0$, i.e. we would like to identify the optimal pair

$$(E_s, \mathbf{s}_s)^{opt} = \arg \max_{E_s > 0, \mathbf{s}_s \in \mathbb{R}^L} E_s \mathbf{s}_s^T \mathbf{R}_{s/s}^{-1} \mathbf{s}_s$$

$$\text{subject to} \quad E_i h_i^2 \mathbf{s}_i^T \mathbf{R}_{p/i}^{-1} \mathbf{s}_i \geq \alpha, \quad i = 1, 2, \ldots, K, \quad (5)$$

$$\mathbf{s}_s^T \mathbf{s}_s = 1, \ E_s \leq E_{max}$$

where E_{max} denotes the maximum available/allowable bit energy for the secondary user.

The optimization task of maximizing a quadratic objective function ($\mathbf{R}_{s/s}^{-1}$ is positive definite) subject to the constraints in (5) is, unfortunately, a non-convex NP-hard (in L) optimization problem [16]. In the following section, we delve into the details of the problem and derive a novel realizable suboptimum solution.

III. PROPOSED COGNITIVE SECONDARY CHANNEL DESIGN

Using the matrix inversion lemma [17] on $\mathbf{R}_{p/i}^{-1}$, we can express the key quadratic constraint expression $\mathbf{s}_i^T \mathbf{R}_{p/i}^{-1} \mathbf{s}_i$ in (5) as

$$\mathbf{s}_i^T \mathbf{R}_{p/i}^{-1} \mathbf{s}_i = \frac{\mathbf{s}_i^T \mathbf{R}_p^{-1} \mathbf{s}_i}{1 - E_i h_i^2 \mathbf{s}_i^T \mathbf{R}_p^{-1} \mathbf{s}_i}, \quad i = 1, 2, \ldots, K, \quad (6)$$

where we recall that $\mathbf{R}_p = E\{\mathbf{r r}^T\}$ is the autocorrelation matrix of the whole input to the primary receiver PR. Then, the PR SINR constraints in (5) become

$$\mathbf{s}_i^T \mathbf{R}_p^{-1} \mathbf{s}_i \geq \frac{\alpha}{E_i h_i^2 + \alpha E_i h_i^2} \triangleq \gamma_i, \quad i = 1, 2, \ldots, K, \quad (7)$$

and the optimization problem can be rewritten as

$$(E_s, \mathbf{s}_s)^{opt} = \arg \max_{E_s > 0, \mathbf{s}_s \in \mathbb{R}^L} E_s \mathbf{s}_s^T \mathbf{R}_{s/s}^{-1} \mathbf{s}_s$$

$$\text{subject to} \quad \mathbf{s}_i^T \mathbf{R}_p^{-1} \mathbf{s}_i \geq \gamma_i, \quad i = 1, 2, \ldots, K, \quad (8)$$

$$\mathbf{s}_s^T \mathbf{s}_s = 1, \ E_s \leq E_{max}.$$

Using the matrix inversion lemma on $\mathbf{R}_p^{-1}$ this time, we see that

$$\mathbf{R}_p^{-1} = \mathbf{R}_{p/s}^{-1} - \frac{E_s h_s^2 \mathbf{R}_{p/s}^{-1} \mathbf{s}_s \mathbf{s}_s^T \mathbf{R}_{p/s}^{-1}}{1 + E_s h_s^2 \mathbf{s}_s^T \mathbf{R}_{p/s}^{-1} \mathbf{s}_s} \quad (9)$$

$$SINR_i = \frac{E\{|\mathbf{w}_{maxSINR,i}^T(\sqrt{E_i}h_i b_i \mathbf{s}_i)|^2\}}{E\{|\mathbf{w}_{maxSINR,i}^T(\sum_{k=1,k\neq i}^K \sqrt{E_k}h_k \mathbf{s}_k b_k + \sqrt{E_s}h_s \mathbf{s}_s b_s + \mathbf{n}_p)|^2\}} = E_i h_i^2 \mathbf{s}_i^T \mathbf{R}_{p/i}^{-1} \mathbf{s}_i, \tag{3}$$

$$SINR_s = \frac{E\{|\mathbf{w}_{maxSINR,s}^T(\sqrt{E_s}q_s b_s \mathbf{s}_s)|^2\}}{E\{|\mathbf{w}_{maxSINR,s}^T(\sum_{k=1}^K \sqrt{E_k}q_k \mathbf{s}_k b_k + \mathbf{n}_s)|^2\}} = E_s q_s^2 \mathbf{s}_s^T \mathbf{R}_{s/s}^{-1} \mathbf{s}_s \tag{4}$$

where $\mathbf{R}_{p/s}$ is the autocorrelation matrix of the input to the primary receiver PR excluding the secondary transmission,

$$\mathbf{R}_{p/s} \triangleq E\{(\sum_{i=1}^K \sqrt{E_i}h_i \mathbf{s}_i b_i + \mathbf{n}_p)(\sum_{i=1}^K \sqrt{E_i}h_i \mathbf{s}_i b_i + \mathbf{n}_p)^T\}$$

$$= \sum_{i=1}^K E_i h_i^2 \mathbf{s}_i \mathbf{s}_i^T + \sigma^2 \mathbf{I}.$$

Then, inserting (9) in (8) we can express the optimization constraints as explicit functions of the code sequence of the secondary user $\mathbf{s}_s$, i.e.

$$\mathbf{s}_i^T \mathbf{R}_{p/s}^{-1} \mathbf{s}_i \geq \frac{E_s h_s^2 \mathbf{s}_i^T \mathbf{R}_{p/s}^{-1} \mathbf{s}_s \mathbf{s}_s^T \mathbf{R}_{p/s}^{-1} \mathbf{s}_i}{1 + E_s h_s^2 \mathbf{s}_s^T \mathbf{R}_{p/s}^{-1} \mathbf{s}_s} + \gamma_i, \ \ i = 1, 2, \ldots, K. \tag{10}$$

For notational simplicity, define the $L \times L$ matrix

$$\mathbf{B}_i \triangleq h_s^2 \mathbf{R}_{p/s}^{-1} \mathbf{s}_i \mathbf{s}_i^T \mathbf{R}_{p/s}^{-1} - \beta_i h_s^2 \mathbf{R}_{p/s}^{-1} \tag{11}$$

where

$$\beta_i \triangleq \mathbf{s}_i^T \mathbf{R}_{p/s}^{-1} \mathbf{s}_i - \gamma_i, \ \ i = 1, 2, \ldots, K. \tag{12}$$

Then, the optimization problem in (8) can be rewritten -for one more time- as

$$\mathbf{x}^{opt} = \arg\max_{\mathbf{x} \in \mathbb{R}^L} \mathbf{x}^T \mathbf{R}_{s/s}^{-1} \mathbf{x}$$

$$\text{subject to} \quad \mathbf{x}^T \mathbf{B}_i \mathbf{x} - \beta_i \leq 0, \quad i = 1, 2, \ldots, K, \tag{13}$$

$$\mathbf{x}^T \mathbf{x} \leq E_{max}$$

where $\mathbf{x}$ is the amplitude-including transmitted signature vector of the secondary user, $\mathbf{x} \triangleq \sqrt{E_s}\mathbf{s}_s$. We notice that for (13) to be solved at the secondary transmitter ST, the primary receiver PR must communicate the matrix parameters $\mathbf{B}_i$ and scalars β_i, $i = 1, 2, \ldots, K$. Therefore, no explicit communication of the primary channel codes and gains is required that may directly compromise the privacy/security of the primary system. In terms of the computational effort, however, (i) $\mathbf{B}_i$, $i = 1, 2, \ldots, K$, are not necessarily positive semidefinite, hence the problem in (13) is in general a non-convex quadratically constrained quadratic program (non-convex QCQP), and (ii) the complexity of a solver of (13) is exponential in the dimension L (NP-hard problem).

To circumvent these two difficulties, we first observe that if we use the trace property of matrices $\mathbf{U}, \mathbf{V}$, $Tr\{\mathbf{UV}\} = Tr(\mathbf{VU})$, we are able to represent the objective function in (13) as

$$\mathbf{x}^T \mathbf{R}_{s/s}^{-1} \mathbf{x} = Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}\} \tag{14}$$

where $\mathbf{X} = \mathbf{x}\mathbf{x}^T$. Thus, the optimization problem in (13) takes the new equivalent matrix form

$$\mathbf{X}^{opt} = \arg\max_{\mathbf{X} \in \mathbb{R}^{L \times L}} Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}\}$$

$$\text{subject to} \quad Tr\{\mathbf{B}_i \mathbf{X}\} \leq \beta_i, \quad i = 1, 2, \ldots, K, \tag{15}$$

$$Tr\{\mathbf{X}\} \leq E_{max}, \ \mathbf{X} \succeq \mathbf{0}, \ rank(\mathbf{X}) = 1$$

where $\mathbf{X} \succeq \mathbf{0}$ denotes that the matrix $\mathbf{X}$ is positive semidefinite.

So far, we have shown that the original secondary cognitive link design problem in (5) is equivalent to the one in (8), (13), and finally (15), and is non-convex NP-hard. To effectively attack the problem anyway, we now propose to *relax the rank constraint* in (15) and proceed by solving the following problem *instead*,

$$\mathbf{X}' = \arg\max_{\mathbf{X} \in \mathbb{R}^{L \times L}} Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}\}$$

$$\text{subject to} \quad Tr\{\mathbf{B}_i \mathbf{X}\} \leq \beta_i, \quad i = 1, 2, \ldots, K, \tag{16}$$

$$Tr\{\mathbf{X}\} \leq E_{max}, \ \mathbf{X} \succeq \mathbf{0}.$$

Then, (16) is a convex polynomial-complexity problem that can be solved using semidefinite programming. Strictly speaking, we can solve (16) in polynomial time within an error $\epsilon > 0$ from its value at the optimum point $\mathbf{X}'$. More specifically, let $f_o \triangleq Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}\}|_{\mathbf{X}=\mathbf{X}'}$, i.e. f_o is the optimum value of the constrained (affine) objective function in (16). Then, for any given $\epsilon > 0$, semidefinite programming guarantees that we can converge in polynomial time (polynomial in the input size L and in the error requirement function $\log 1/\epsilon$) to a solution that lies in $(f_o - \epsilon, f_o)$ [18]. In this paper, for the semidefinite programming problem in (16), we propose to use a primal-dual interior-point method [19]. In particular, we consider the problem in (16) as the primal optimization problem, we create a differently parameterized equivalent dual problem, and then solve both problems iteratively in a coupled fashion. Then, each iteration can be implemented in $O(L^3)$ and the algorithm converges after $O(L \log 1/\epsilon)$ iterations to the matrix $\mathbf{X}''$ that makes the objective function $Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}\}$ attain a value within $(f_o - \epsilon, f_o)$. The proposed method is outlined in Fig. 2. We note that relaxing the rank constraint of the non-convex NP-hard problem in (15) we created the convex optimization problem in (16) that can be solved in $O(L^4 \log 1/\epsilon)$ time (by semidefinite programming methods as described in Fig. 2). Of course, because of the constraint relaxation itself the objective function evaluated at the optimum point $\mathbf{X}'$ in (16) is just an upper bound on the value of the objective function evaluated at the optimum point of interest $\mathbf{X}^{opt}$ of (15), $Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}^{opt}\} \leq Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}'\}$. Moreover, $\mathbf{X}'$ is not available exactly either and instead we have $\mathbf{X}''$ with $Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}''\} \in (Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}'\} - \epsilon, Tr\{\mathbf{R}_{s/s}^{-1}\mathbf{X}'\})$.

To summarize our developments so far for the cognitive design of a code-division secondary link, first, for the given primary SINR-QoS threshold $\alpha > 0$, we test whether β_i, $i = 1, 2, \ldots K$, in (12) are all greater than zero. If this is not true, then the SINR-QoS constraints for the primary users cannot be met and outright no secondary transmission is allowed (see flow-chart in Fig. 3). Otherwise, we run the procedure of Fig. 2 which returns matrix $\mathbf{X}''$. If the rank of $\mathbf{X}''$ is 1 with eigenvalue, eigenvector pair $\lambda_1, \mathbf{a}_1$, then we already
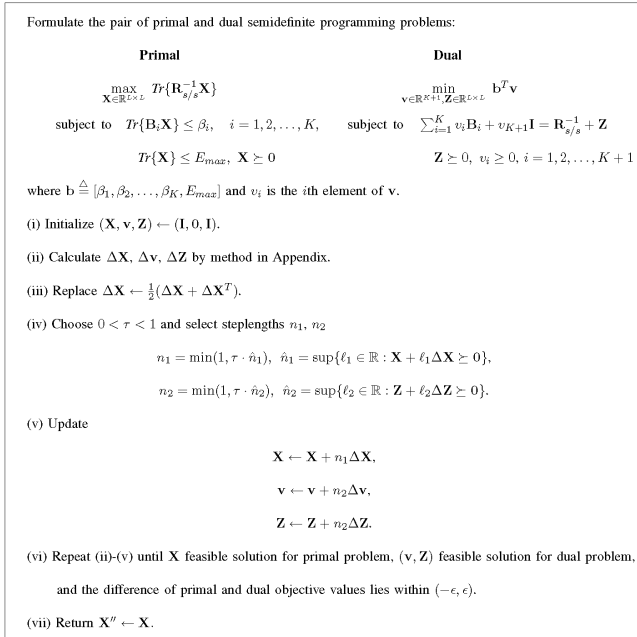
Fig. 2. Proposed interior-point algorithm for solving (16).

Fig. 3. Flow-chart of proposed power and code allocation algorithm for secondary link.

have our secondary link design with signature $\mathbf{s}_s = \mathbf{a}_1$ and transmission amplitude $E_s = \lambda_1$. If the rank of $\mathbf{X}''$ is not 1, further work is needed as described below.

When $\mathbf{X}'$ of (16) (or in practice $\mathbf{X}''$ returned by Fig. 2) happens to be of rank 1 with eigenvalue, eigenvector pair $\lambda_1, \mathbf{a}_1$, then $\mathbf{X}' \equiv \mathbf{X}^{opt}$ in (15) and $\mathbf{x}^{opt} = \sqrt{\lambda_1}\mathbf{a}_1$ in (13). Otherwise, there is no direct path from $\mathbf{X}'$ of (16) to $\mathbf{x}^{opt}$ in (13). In this case, we may simply consider changing the search for an optimal vector in (13) to a search for an optimal probability density function (pdf) of vectors that maximizes the average objective function subject to average constraints, i.e.

$$f^{opt}(\mathbf{x}) = \arg\max_{f(\mathbf{x})} \ E\{\mathbf{x}^T \mathbf{R}_{s/s}^{-1} \mathbf{x}\}$$
$$\text{subject to} \quad E\{\mathbf{x}^T \mathbf{B}_i \mathbf{x}\} \leq \beta_i, \quad i = 1, 2, \ldots, K, \quad (17)$$
$$E\{\mathbf{x}^T \mathbf{x}\} \leq E_{max}$$

where $f(\mathbf{x})$ denotes the probability density function of $\mathbf{x}$. This switch to a statistical optimization problem has been known as the "randomized method" in semidefinite programming literature [18]. Using the commutative property between trace and expectation operators, the pdf optimization problem in (17) takes the equivalent form

$$f^{opt}(\mathbf{x}) = \arg\max_{f(\mathbf{x})} \ Tr\{\mathbf{R}_{s/s}^{-1} E\{\mathbf{x}\mathbf{x}^T\}\}$$
$$\text{subject to} \quad Tr\{\mathbf{B}_i E\{\mathbf{x}\mathbf{x}^T\}\} \leq \beta_i, \quad i = 1, 2, \ldots, K, \quad (18)$$
$$Tr\{E\{\mathbf{x}\mathbf{x}^T\}\} \leq E_{max}.$$

We can show that $f^{opt}(\mathbf{x})$ is in fact Gaussian with $\mathbf{0}$ mean and covariance matrix $\mathbf{X}'$, $f^{opt}(\mathbf{x}) = \mathcal{N}(\mathbf{0}, \mathbf{X}')$. With $\mathbf{X}''$ from Fig. 2 as a close approximation of $\mathbf{X}'$, we can draw now a sequence of samples $\mathbf{x}_1, \mathbf{x}_2, \ldots, \mathbf{x}_P$ from $\mathcal{N}(\mathbf{0}, \mathbf{X}'')$. We test all of them for "feasibility" on the constraints of (13) whether $\mathbf{x}_p^T \mathbf{B}_i \mathbf{x}_p \leq \beta_i, \ \forall i = 1, 2, \ldots, K,$ and $\mathbf{x}_p^T \mathbf{x}_p \leq E_{max}, \ p = 1, 2, \ldots, P,$ and among the feasible vectors (if any) we choose
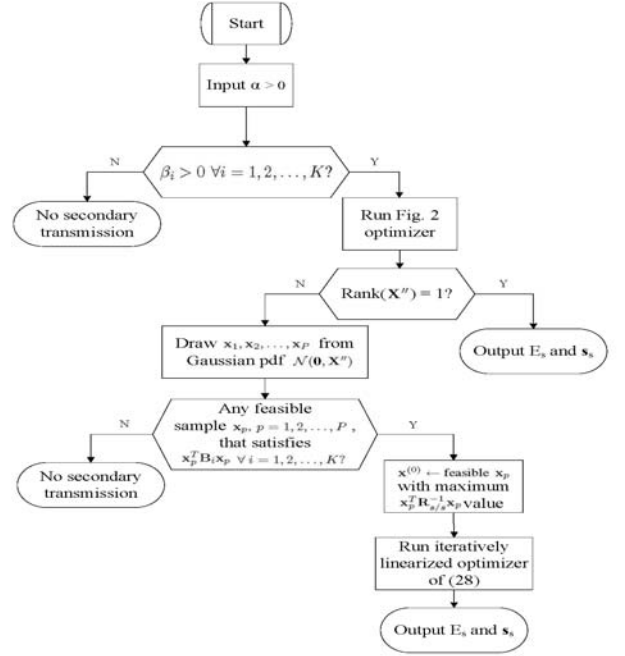
the one, say $\mathbf{x}^{(0)}$, with maximum $\mathbf{x}^T \mathbf{R}_{s/s}^{-1} \mathbf{x}$ objective function value (see flow-chart in Fig. 3). We could have suggested at this time a cognitive secondary link design with $\sqrt{E_s}\mathbf{s}_s = \mathbf{x}^{(0)}$. Instead, we will use $\mathbf{x}^{(0)}$ as an initialization point to an iterative procedure that will lead to a much improved link design vector. The iterative procedure is developed below and its performance is evaluated by simulation studies in the next section.

First we express $\mathbf{R}_{s/s}$ as

$$\mathbf{R}_{s/s} = \mathbf{S}\Sigma\mathbf{S}^T + \sigma^2\mathbf{I} \quad (19)$$

where $\mathbf{S} \triangleq [\mathbf{s}_1, \mathbf{s}_2, \ldots, \mathbf{s}_K]$ denotes the matrix with columns the signatures of the primary users, and $\Sigma = \text{diag}(E_1 q_1^2, E_2 q_2^2, \ldots, E_K q_K^2)$. Using the matrix inversion lemma,

$$\mathbf{R}_{s/s}^{-1} = \frac{1}{\sigma^2}\mathbf{I} - \frac{1}{\sigma^4}\mathbf{S}(\Sigma^{-1} + \frac{1}{\sigma^2}\mathbf{S}^T\mathbf{S})^{-1}\mathbf{S}^T. \quad (20)$$

Substitution of (20) in the objective function of (13) leads to

$$\mathbf{x}^T \mathbf{R}_{s/s}^{-1} \mathbf{x} = \frac{1}{\sigma^2}\mathbf{x}^T\mathbf{x} - \frac{1}{\sigma^4}\mathbf{x}^T\mathbf{Q}\mathbf{x} \quad (21)$$

where $\mathbf{Q} \triangleq \mathbf{S}(\Sigma^{-1} + \frac{1}{\sigma^2}\mathbf{S}^T\mathbf{S})^{-1}\mathbf{S}^T$. In (21), the first term $\frac{1}{\sigma^2}\mathbf{x}^T\mathbf{x}$ is a convex function while the second term $-\frac{1}{\sigma^4}\mathbf{x}^T\mathbf{Q}\mathbf{x}$ is a concave function (which implies that $\frac{1}{\sigma^4}\mathbf{x}^T\mathbf{Q}\mathbf{x}$ is convex). Based on the first-order conditions of convex functions [20], we have

$$\mathbf{x}^T\mathbf{x} \geq 2\mathbf{x}^{(0)T}\mathbf{x} - \mathbf{x}^{(0)T}\mathbf{x}^{(0)} \quad (22)$$

where $\mathbf{x}^{(0)}$ denotes an initial feasible vector. Then, we combine (21) and (22) and form an optimization problem that maximizes the following concave function

$$\frac{2}{\sigma^2}\mathbf{x}^{(0)T}\mathbf{x} - \frac{1}{\sigma^4}\mathbf{x}^T\mathbf{Q}\mathbf{x} - \frac{1}{\sigma^2}\mathbf{x}^{(0)T}\mathbf{x}^{(0)} \quad (23)$$

that leads to a *suboptimum* solution for our original problem in (13). To maximize (23) in view of our constraints in (13), we restrict all non-convex constraints into convex sets (linearization). In particular, we consider the non-convex constraints

$$\mathbf{x}^T \mathbf{B}_i \mathbf{x} - \beta_i \leq 0, \quad i \in \mathcal{I}_{nc}, \tag{24}$$

where $\mathcal{I}_{nc}$ denotes the set of all indices $i \in \{1, 2, \ldots, K\}$ for which $\mathbf{x}^T \mathbf{B}_i \mathbf{x}$ is a non-convex function. Then, we decompose the matrix $\mathbf{B}_i$ into its positive and negative parts

$$\mathbf{B}_i = \mathbf{B}_i^+ - \mathbf{B}_i^- \tag{25}$$

where $\mathbf{B}_i^+ = h_s^2 \mathbf{R}_{p/s}^{-1} \mathbf{s}_i \mathbf{s}_i^T \mathbf{R}_{p/s}^{-1}$ and $\mathbf{B}_i^- = \beta_i h_s^2 \mathbf{R}_{p/s}^{-1}$ are all positive semidefinite. Therefore, the original constraints (24) can be written as

$$\mathbf{x}^T \mathbf{B}_i^+ \mathbf{x} - \beta_i \leq \mathbf{x}^T \mathbf{B}_i^- \mathbf{x}, \quad i \in \mathcal{I}_{nc}, \tag{26}$$

where both sides of the inequality are convex quadratic functions. Linearization of the right-hand side of (26) around the vector $\mathbf{x}^{(0)}$ leads to

$$\mathbf{x}^T \mathbf{B}_i^+ \mathbf{x} - \beta_i \leq \mathbf{x}^{(0)T} \mathbf{B}_i^- \mathbf{x}^{(0)} + 2\mathbf{x}^{(0)T} \mathbf{B}_i^- (\mathbf{x} - \mathbf{x}^{(0)}), i \in \mathcal{I}_{nc}. \tag{27}$$

In (27), the right-hand side is an affine lower bound on the original function $\mathbf{x}^T \mathbf{B}_i^- \mathbf{x}$. It is thus implied that the resulting constraints are convex and more conservative than the original ones, hence the feasible set of the linearized problem is a convex subset of the original feasible set. Thus, by linearizing the concave parts of all constraints, we obtain a set of convex constraints that are tighter than the original non-convex ones. Now, the original optimization problem takes the form

$$\mathbf{x}^{(1)} = \arg \max_{\mathbf{x} \in \mathbb{R}^L} \frac{2}{\sigma^2} \mathbf{x}^{(0)T} \mathbf{x} - \frac{1}{\sigma^4} \mathbf{x}^T \mathbf{Q} \mathbf{x} - \frac{1}{\sigma^2} \mathbf{x}^{(0)T} \mathbf{x}^{(0)}$$

$$\text{subject to } \mathbf{x}^T \mathbf{B}_i^+ \mathbf{x} - \mathbf{x}^{(0)T} \mathbf{B}_i^- (2\mathbf{x} - \mathbf{x}^{(0)}) - \beta_i \leq 0, \ i \in \mathcal{I}_{nc},$$

$$\mathbf{x}^T \mathbf{B}_i \mathbf{x} - \beta_i \leq 0, \quad i \in \overline{\mathcal{I}}_{nc}, \tag{28}$$

$$\mathbf{x}^T \mathbf{x} \leq E_{max}$$

where $\overline{\mathcal{I}}_{nc} = \{1, 2, \ldots, K\} - \mathcal{I}_{nc}$. The problem in (28) is a convex QCQP problem and can be solved efficiently by standard convex system solvers [21] to produce a new feasible vector $\mathbf{x}^{(1)}$. The objective function $\mathbf{x}^T \mathbf{R}_{s/s}^{-1} \mathbf{x}$ in (13) evaluated at $\mathbf{x}^{(1)}$ takes a value that is larger than or equal to its value at $\mathbf{x}^{(0)}$. Repeating iteratively the linearization procedure, we can obtain a sequence of feasible vectors $\mathbf{x}^{(0)}$, $\mathbf{x}^{(1)}$, $\mathbf{x}^{(2)}, \ldots, \mathbf{x}^{(T)}$ with non-decreasing values of the objective function in (13). This procedure converges after very few (eight or nine) iterations as demonstrated experimentally in the following section.

IV. SIMULATION STUDIES

We consider a primary CDMA system with signature length (system processing gain) $L = 16$ and K synchronous users. We are interested in establishing a secondary code-division transmitter/receiver pair when the primary system is fully loaded to overloaded, say K varies from 16 to 20. All signatures for primary users are generated from a minimum total-squared-correlation optimal binary signature set which achieves the Karystinos-Pados (KP) bound for each (K, L) pair of values[3] [22]-[24]. The transmission SNRs of the K

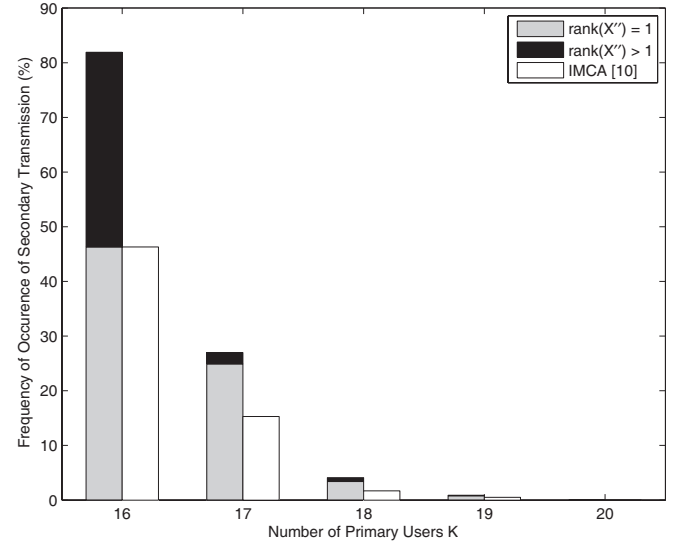[3] For $L = 16$, when $K \leq L$ the KP-optimal sequences coincide with the familiar Walsh-Hadamard signature codes.

Fig. 4. Secondary transmission percentage as a function of the number of primary users under Cases $rank(\mathbf{X}'') = 1$ and > 1 (the study includes also the code assignment scheme in [10]).

primary users are all set equal to $\frac{E_i}{\sigma^2} = 15dB$, $i = 1, 2, \ldots K$; the maximum allowable transmission SNR for the secondary link is set to $\frac{E_{max}}{\sigma^2} = 12dB$. The channel coefficients h_i and q_i, $i = 1, 2, \ldots, K$, (see Fig. 1) are taken to be the magnitude of independant complex Gaussian random variables with mean 0 and variance 4; the same holds true for h_s and q_s. The receiver SINR threshold for primary users is set to $\alpha = 2dB$ which corresponds to an average raw bit-error-rate (BER) at the output of the maximum SINR linear filter receiver of about 10^{-1}. Ten thousand (10,000) system/secondary-line optimization experiments are run under the described (quasi-static) flat fading conditions. When random vector drawing is necessitated by the flow-chart in Fig. 3, $P = 50$ test vector points are generated.

In Fig. 4, we plot as a function of the number of primary system users K the percentage of time that secondary transmission is enabled directly under the case $rank(\mathbf{X}'') = 1$ or by the iterative linearized optimizer as well as the "Interference-Minimizing-Code-Assignment" (IMCA) scheme in [10]. We observe that significant opportunity exists for cognitive secondary transmission when the primary system is fully loaded ($K = L = 16$). As we expect, Fig. 4 shows that the frequency of secondary transmissions reduces as the primary system load increases. We observe also that our proposed scheme offers more opportunities for cognitive secondary transmission than [10].

In Fig. 5, we test the quality of the secondary transmission line (the pre-set SINR-QoS of the primary system is -of course- guaranteed by the algorithmic procedure) and the significance of the iterative linearized optimizer in (28) (see flow-chart in Fig. 3). We fix the primary system load $K = 16$ (fully loaded) and plot the secondary receiver *average* SINR for the experimental instants of $rank(\mathbf{X}'') > 1$ as a function of the iteration of the optimizer initialized at the point/design $\mathbf{x}^{(0)}$ that is the best out of $P = 50$ samples drawn from the $\mathcal{N}(\mathbf{0}, \mathbf{X}'')$ pdf. It is pleasing to observe that eight or nine iterations are enough for effective convergence.
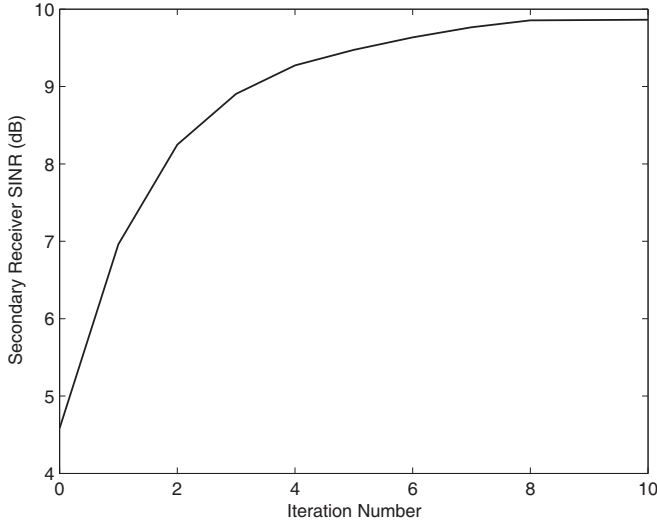
Fig. 5. Secondary receiver SINR as a function of the iteration step of the linearized optimizer in (28) initialized at the best feasible sample out of $P = 50$ drawings from the $\mathcal{N}(\mathbf{0}, \mathbf{X}'')$ pdf.

Fig. 6. Instantaneous output SINR of a primary signal against SINR-QoS threshold α (thick line) and instantaneous output SINR of secondary signal.

Finally, in Fig. 6, to gain visual insight into the operation of the primary/secondary system we plot the *instantaneous* receiver SINR of a primary signal and the secondary signal for the case $K = 17$ over an experimental data record sequence of 1000 Rayleigh fading channel realizations. Missing secondary signal SINR values indicate the instances when no secondary transmission was allowed. The proposed scheme almost doubled the occurrences of secondary transmission compared to [10]. When secondary transmissions do occur for both schemes, the joint power and sequence optimization executed by the proposed scheme results in superior SINR performance for the secondary receiver over [10].

V. CONCLUSIONS

We considered the general problem of establishing a secondary code-division line alongside a primary code-division multiple-access system. We formulated the problem as the search for the secondary amplitude, code transmission pair that maximizes the secondary line output SINR subject to the condition that all primary signal output SINR values are maintained above a given SINR-QoS threshold value. Regretfully, in a common Rayleigh fading wireless environment, the formulated constrained optimization problem is non-convex and NP-hard in the code vector dimension.

Nevertheless, in pursuit of a computationally manageable and performance-wise appealing suboptimal solution, we first converted the amplitude/code-vector optimization problem to an equivalent matrix optimization problem under a rank-1 constraint. Disregarding ("relaxing" in formal language) the rank-1 constraint makes the problem amenable to an "easy" polynomial-cost semidefinite programming solution. When luckily, a rank-1 matrix happens to be returned, optimal secondary-line design is achieved. For the common case of a higher rank, an iterative linearized polynomial-cost convex optimizer is developed with much appealing (yet suboptimal) amplitude/code-vector design solutions after a few iterations.

Cognitive code-division radios combine in principle the bandwidth efficiency characteristics of cognitive operation and
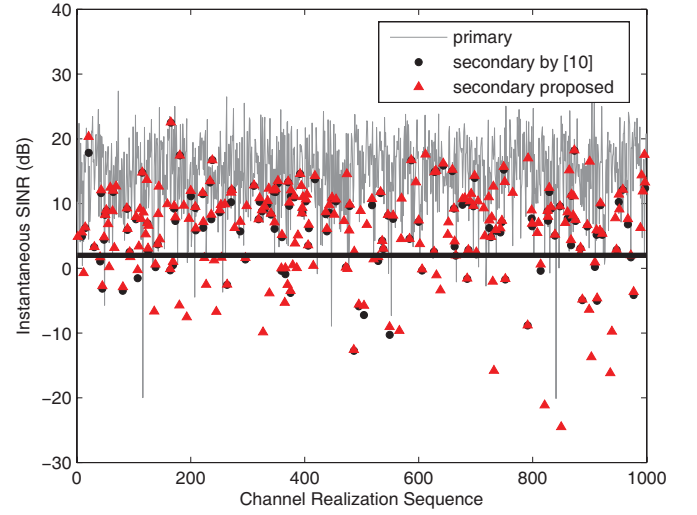
code-division multiple accessing and are expected to find a place in future communication systems. To that extend, the developments presented in this paper constitute an early contribution that can be helpful in benchmarking future efforts.

APPENDIX A
CALCULATION OF $\Delta\mathbf{X}$, $\Delta\mathbf{v}$, $\Delta\mathbf{Z}$ OF FIG. 2

Let $vec\{\cdot\}$ denote column-by-column matrix vectorization and $mat\{\cdot\}$ the exact inverse operation. Choose $0 \leq \delta < 1$ and define $\mu \triangleq \delta\frac{Tr\{\mathbf{XZ}\}}{L}$.

In Fig. 2, $\Delta\mathbf{X}$, $\Delta\mathbf{v}$, $\Delta\mathbf{Z}$ are obtained by solving the following linear system

$$\begin{bmatrix} \mathbf{0} & \mathbf{B}^T & \mathbf{I} \\ \mathbf{B} & \mathbf{0} & \mathbf{0} \\ \mathbf{E} & \mathbf{0} & \mathbf{F} \end{bmatrix} \begin{bmatrix} vec\{\Delta\mathbf{X}\} \\ \Delta\mathbf{v} \\ vec\{\Delta\mathbf{Z}\} \end{bmatrix} = \begin{bmatrix} vec\{\mathbf{T}_1\} \\ \mathbf{t}_2 \\ vec\{\mathbf{T}_3\} \end{bmatrix} \quad (29)$$

where

$$\mathbf{B}_{(K+1)\times L^2} \triangleq \begin{bmatrix} (vec\{\mathbf{B}_1\})^T \\ \vdots \\ (vec\{\mathbf{B}_K\})^T \\ (vec\{\mathbf{I}\})^T \end{bmatrix},$$

$\mathbf{T}_{1\,L\times L} \triangleq \mathbf{R}_{s/s}^{-1} + \mathbf{Z} - mat\{\mathbf{B}^T\mathbf{v}\}$, $\mathbf{t}_{2\,(K+1)\times 1} \triangleq \mathbf{b} - \mathbf{B}vec\{\mathbf{X}\}$, $\mathbf{T}_{3\,L\times L} \triangleq \mu\mathbf{I} - \mathbf{XZ}$, $\mathbf{E}_{L^2\times L^2} \triangleq \mathbf{Z} \otimes \mathbf{I}$, and $\mathbf{F}_{L^2\times L^2} \triangleq \mathbf{I} \otimes \mathbf{X}$ ($\otimes$ denotes the standard Kronecker product). Applying Gauss elimination, the solution is

$$\Delta\mathbf{v} = (\mathbf{B}\mathbf{E}^{-1}\mathbf{F}\mathbf{B})^{-1}(\mathbf{t}_2 + \mathbf{B}\mathbf{E}^{-1}(\mathbf{F}vec\{\mathbf{T}_1\} - vec\{\mathbf{T}_3\})), \quad (30)$$

$$vec\{\Delta\mathbf{X}\} = -\mathbf{E}^{-1}(\mathbf{F}(vec\{\mathbf{T}_1\} - \mathbf{B}^T\Delta\mathbf{v}) - vec\{\mathbf{T}_3\}), \quad (31)$$

$$vec\{\Delta\mathbf{Z}\} = vec\{\mathbf{T}_1\} - \mathbf{B}^T\Delta\mathbf{v}. \quad (32)$$

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